LEVEL II

ELECTRONICALLY TUNED UHF RECEIVER PRESELECTOR

E-Systems

Thomas Swanson

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SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

REPORT DOCUMENTATION PAGE	READ INSTRUCTIONS BEFORE COMPLETING FORM

1. REPORT NUMBER	2. GOVT ACCESSION NO.	3. RECIPIENT'S CATALOG NUMBER
RADC-TR-79-210		

4. TITLE (and Subtitle)	5. TYPE OF REPORT & PERIOD COVERED
ELECTRONICALLY TUNED UHF RECEIVER PRESELECTOR	Final Technical Report 27 Oct 77 — 30 Oct 78
	6. PERFORMING ORG. REPORT NUMBER
	N/A

7. AUTHOR(s)	8. CONTRACT OR GRANT NUMBER(s)
Thomas Swanson	F30602-77-C-0200

9. PERFORMING ORGANIZATION NAME AND ADDRESS	10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS
E-Systems, ECI Division 1501-72nd Street, North St. Petersburg FL 33733	62702F 23380402

11. CONTROLLING OFFICE NAME AND ADDRESS	12. REPORT DATE
Rome Air Development Center (RBCT) Griffiss AFB NY 13441	August 1979
	13. NUMBER OF PAGES
	102

14. MONITORING AGENCY NAME & ADDRESS(if different from Controlling Office)	15. SECURITY CLASS. (of this report)
Same	UNCLASSIFIED
	15a. DECLASSIFICATION/DOWNGRADING SCHEDULE
	N/A

16. DISTRIBUTION STATEMENT (of this Report)

Approved for public release; distribution unlimited

17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)

Same

18. SUPPLEMENTARY NOTES

RADC Project Engineer: Daniel E. Warren (RBCT)

19. KEY WORDS (Continue on reverse side if necessary and identify by block number)

Electromagnetic Compatibility   Electronic Tuning
Communications
Filters
Interference
Intermodulation

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

This report describes design and construction of a prototype improved electronically tunable UHF receiver preselector for interference reduction in Air Force communication systems. This research work was directed toward developing the technology for frequency agile preselectors to protect solid state receivers from high level signals originating from colocated UHF transmitters. The design approach consists of tuning a modified comb-line filter structure with high Q UHF capacitors and pin diode switches. Equations and design considerations are given for construction of tuning networks which have

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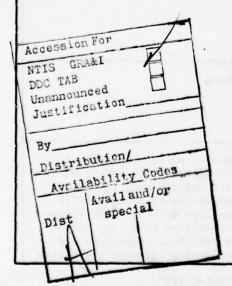SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

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(to the N-th power)

$Cx2^N$ binary capacitance steps. A computer program was used to predict the filters performance based on the values of stray capacitance and inductance of the tuning networks. The tuning networks were then optimized which ensured that spurious resonances fell well outside of the tuning range of the filters. The filters were built based on these computer modeled results.

Test data showed input VSWR variation versus the tuning frequency of the filters. These changes were found to be caused by variation in current loading at the tap point of the resonator resulting from the different values of capacitance selected in the tuning network. Improved filter performance was obtained by grounding the input transformers in the vicinity of the tuning network tap point.

A computer controlled test set-up was used for electrical performance testing over the 225 to 400 MHz frequency range. The completed preselector had an average 3 dB bandwidth of 1.8 MHz. In the center of the UHF band, the preselectors noise figure was such to degrade a typical 14 dB noise figure receiver by 1 dB and it had approximately unity gain, although the noise figure degraded at the upper and lower edges of the UHF band. The intercept point for tones greater than 1.5 MHz from fc was greater than +40 dBm and for tones greater than 10 MHz from fc was greater than +70 dBm.

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EVALUATION

This report documents the efforts made to advance the state-of-the-art with respect to electronically tuned UHF receiver preselectors. The goal was to suppress the effects of high power interfering signals from colocated transmitters, without seriously degrading the sensitivity, tuning speed or compactness of recently developed solid state transceivers. Emphasis is placed upon tuning range (225-400 MHz), compactness (2 pounds and 60 cubic inches), low insertion loss (.5 dB) and narrow bandwidth (500 KHz).

These goals were partially met, but such a filter alone will not solve the interference problems. It is possible to expand this capability somewhat by extending the effort, but it is not likely that the state-of-the-art will be advanced enough to completely solve the interference problem. An approach more likely to succeed is to incorporate the results of this effort with those of another effort which emphasize expanding the dynamic range of receiver front ends. Together these two efforts may solve most of the UHF interference problems.

This effort was in support of TPO R4C.

Daniel E. Warren

DANIEL E. WARREN
Project Engineer

I INTRODUCTION

A. Background

The objective of this contract was to develop an
electronically tunable UHF preselector to reduce commonly en-
countered collocation interference where high power (1000 watt)
transmitters are required to operate simultaneously near sensi-
tive UHF receivers. Recently this problem has been aggravated
by the use of solid state receiver front end circuits. These
receivers are desensitized by signal levels of -10 dBm or above.
Therefore, the purpose of the preselector is to attenuate sig-
nals from these collocated transmitters which can be as high
as +20 dBm at the receiver input to levels below the desensi-
tization threshold.

This contract has relied heavily on data from three
previous RADC contracts which involved work on UHF narrowband
pin diode switched filters. These contracts advanced the
"state of art" and proved that low loss filters can be imple-
mented tuning 12 to 18% portions of the UHF band. The prev-
ious technique was not chosen, however, due to its relatively
narrow tuning range since four filters are required to cover
the UHF band and they are single pole devices.

This contract requires a 3 to 4 pole filter selectivity
depending on allowances for design margin and resonator

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tracking inaccuracy. Using the previous technique, 16 single
pole filters would be necessary occupying approximately 800
cubic inches of space. A new technique was chosen, heretofore
untested as a receiver preselector. It uses pin diode switched
capacitors to tune a 2 pole comb line filter over the entire
UHF band. The preselector's size is reduced by a factor of
10 to only 80 cubic inches. The final preselector was imple-
mented as two 2 pole filters separated by a high dynamic range,
low noise figure amplifier. The preselector's input filter
has the most critical electrical requirements. It was de-
signed to occupy 51 cubic inches where the less critical second
filter is much smaller, packaged in only 26 cubic inches.
The comb line technique has the following advantages and dis-
advantages:

Advantages   • Each filter tunes the entire UHF band.

                • It maintains constant resonator to resonator
                  coupling across the UHF band.

                • It has a high packaging efficiency, in other
                  words high Q_u for such a small size.

Disadvantages   • It requires careful construction since close
                  resonator proximity can cause tuning inter-
                  action.

                • Its small size makes the filter practically a
                  lumped constant network requiring lower loss
                  pin diodes and capacitors than substantially
                  larger filter networks.

B.   The Design Objectives

4.1.1	Frequency range	225-400 MHz
4.1.2	3 dB bandwidth	500 kHz*
4.1.3	30 dB bandwidth	3 MHz
4.1.4	Out-of-band rejection	40 dB
4.1.5	Stability	$\pm$5 kHz**
4.1.6	Noise figure	Will not degrade a 14 dB NF amplifier to no more than 15 dB NF
4.1.7	Insertion loss	Less than 0.5 dB
4.1.8	Maximum out-of-band signal level	+20 dBm
4.1.9	3rd Order inter-modulation products	Less than -30 dBm resulting from 2 +20 dBm inputs
4.1.10	Tuning ability	Compatible with AN/ARC-164, AN/ARC-171
4.1.11	Tuning speed	10 milliseconds
4.1.12	Size	60 cubic inches
4.1.13	Weight	2 pounds

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* Originally 300 kHz but changed to 500 kHz at first technical interchange meeting.

** Sufficient stability so that the performance of an ARC-164 is not degraded.

C.   Summary

During this contract a high dynamic range UHF pre-
selector was developed which tunes from 225 to 400 MHz. A
technique was successfully tested which maintained binary
tuning increments over this relatively wide frequency range.
Two two-pole filters were designed which occupied a total
volume of only 80 cubic inches, approximately one-tenth that
of earlier techniques.

We found that even with a small volume filter such as
this the major losses in the tuning network were due to the
PIN diode switches and not the tuning capacitors or cavity
structure itself. This is explained in detail in Section 4.
The bandwidth and insertion loss goals were not met due to
the losses occurring in the PIN diodes even though the diodes
selected have the highest Q which are usable in the UHF band.

Our measured data on a number of diodes indicates losses
can be reduced by optimizing the diode package and mounting
technique, although time did not permit any study of this
under this contract. Also further study of the loss phenomena
in the silicon junction is certainly appropriate.

Although the majority of the effort of this contract
was devoted toward optimizing this type of filter structure

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we feel we are just beginning to see the potential performance capability of these PIN diode tuned designs.

It is felt further progress can be made if the work is directed toward two main areas:

1. Developing lower loss PIN diodes for use in the UHF band.

2. Further investigating interresonator coupling struc- tures which maintain constant coupling versus changes in capacitive loading from tuning networks.

Figure 1.1 is a summary table listing the performance of the preselector versus design goals. The measured performance of the preselector is listed in detail in Section 7.

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GOALS

	Met	Partly Met	Did Not Meet
Frequency Range	X		
3 dB Bandwidth			X
40 dB Bandwidth			X
Out-of-Band Rejection	X		
Stability	X		
Noise Figure		X	
Insertion Loss		X	
Maximum Out-of-Band Signal Level	X		
3rd Order Inter-modulation Products		X	
Tuning Ability	X		
Tuning Speed	X		
Size			X
Weight			X

Figure 1.1   Design Goals Versus Compliance

D.   Accomplishments

* Did preselector tradeoff analysis to achieve lowest unloaded Q filter configuration.

* Single filter tuning from 225 to 400 MHz.

* UHF tuning increments of less than 250 kHz.

* Achieved a filter intercept point of approximately +75 dBm.

* Improved upon electronically tuned pin diode switched filter computer modeling techniques.

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*   Developed a technique of binary tuning increments, minimizing the number of pin diode bits and implemented this in a comb line structure.

*   Developed and implemented a low loss pin diode biasing technique which minimizes frequency mistuning.

*   Developed pin diode driving circuitry handling up to 2 amps forward current and 500 V reverse bias and interfaced this to the standard IEEE interface bus for operation with an automatic network analyzer.

*   Achieved a tuning speed of less than one-half millisecond which could be improved to less than 5 μsec. with increased pin diode driver complexity.

*   Developed an improved input transformer which provides more uniform coupling versus frequency.

*   Developed a method to shorten the normal length of comb line structures.

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II THE PRESELECTOR'S FILTER REQUIREMENTS

A.   <u>General</u>

The design of the preselector consists of cascaded filter
and amplifier modules. This section describes how the speci-
fications for each filter and amplifier within the preselector
can be derived from the goals of the overall preselector.
Some of these goals are interrelated with one another and the
tradeoffs between the amount of filter attenuation and ampli-
fier gain, etc., become rather complex. This section is orga-
nized so that goals which are not interrelated are discussed
first since they directly define various filter and amplifier
characteristics.

In this section, we will make the following assumptions:

1.   All filters whether single pole or double pole use
the same types of components, therefore have the same unloaded
Q.

2.   The amplifiers will be off-the-shelf components since
this contract's engineering effort is concentrated toward the
design of small electronically tunable filters.

This section's objective is to describe the implementation
of filters and amplifiers which require the least stringent
unloaded Q, yet still meet all performance goals.

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B.   <u>The Noise Figure Determines the Insertion Loss</u>

The noise figure of the preselector receiver system is a function of preselector gain, preselector noise figure, and receiver noise figure. The block diagram of the receiver system is shown in Figure 2.1.
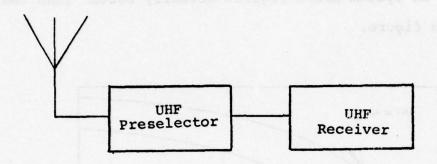
Figure 2.1 Receiver Preselector System

The preselector's noise figure objective is less than 15 dB, assuming the receiver has a 14 dB noise figure like a typical airborne receiver. The noise figure is calculated as follows,

$$F_{TOT} = F_1 + \frac{F_2 - 1}{G_1}$$

where, F_{TOT} = the total noise figure,

F_1 = the noise figure of the preselector,

F_2 = the receiver's noise figure, and

G_1 = the preselector's gain.

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The curve shown in Figure 2.2 is a plot of this equation. Points falling below the 15 dB noise figure line would be within the specification, while those above the line would be out of the specification. The additional 14 dB noise figure curve plotted for information only shows the gain and noise figure which result in system noise figures actually better than the radio's noise figure.
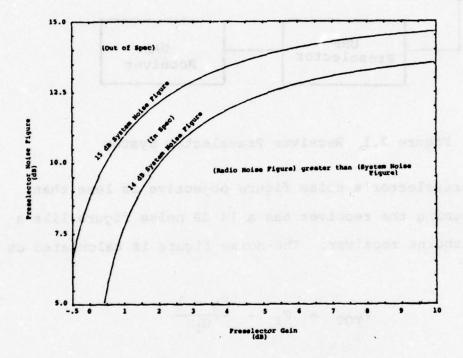
Figure 2.2   Preselector Noise Figure
Versus Preselector Gain
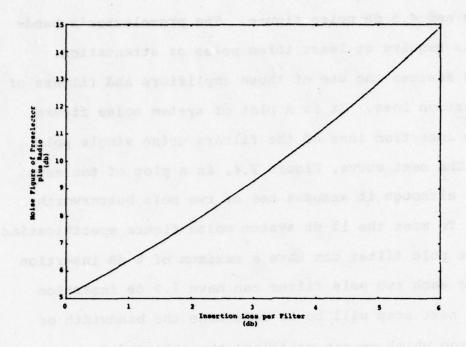
A study of off-the-shelf amplifiers indicates that the Anzac AM-101 has a higher intercept point and a lower noise figure than any others which are readily available. The

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manufacturer's test data sheet shows typical performance as 10 dB gain and 4.5 dB noise figure. The preselector's bandwidth goals require at least three poles of attenuation. Figure 2.3 assumes the use of these amplifiers and filters of equal insertion loss. It is a plot of system noise figure versus the insertion loss of the filters using single pole filters. The next curve, Figure 2.4, is a plot of the same parameters although it assumes use of two pole Butterworth filters. To meet the 15 dB system noise figure specification, each single pole filter can have a maximum of 6 dB insertion loss, while each two pole filter can have 7.5 dB insertion loss. The next step will be to determine the bandwidth or loaded Q from which we can calculate the unloaded Q.

C.   The Loaded Q Calculations

To find the individual bandwidths of each filter, we need to know the number of filters in cascade and the type of each filter. Naturally, as in the first case, the 3 dB bandwidth of the total of three filters in cascade is equal to the 1 dB bandwidth of each filter; and, for the second case, the 3 dB bandwidth of two filters in cascade is equal to the 1.5 dB bandwidth of each filter.

For the preselector shown in Figure 2.3, using single pole filters, the 1 dB bandwidth of each filter will be 500 kHz; although in order to calculate the loaded Q, the 3 dB bandwidth is required. From reference 4, the 1 dB to 3 dB

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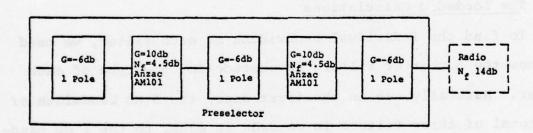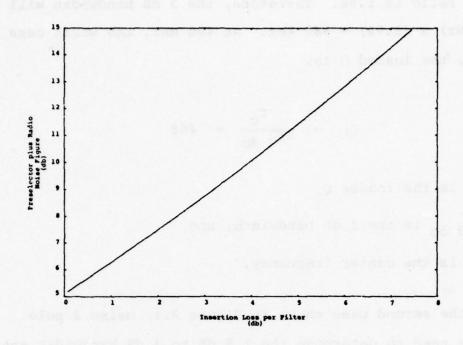
Figure 2.3 Noise Figure of Preselector
and Radio System Versus In-
sertion Loss Per Filter (Case 1)

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Preselector plus Radio
Noise Figure
(db)

Insertion Loss per Filter
(db)

| Preselector | | | |
| G=-7.5db $I_p>60$ 2^PPole | G=10db $N_f=4.5db$ $I_p=28dbm$ | G=-7.5db $I_p>60$ 2^PPole | G=10db $N_f=4.5db$ $I_p=28dbm$ | Radio N_f+14db |
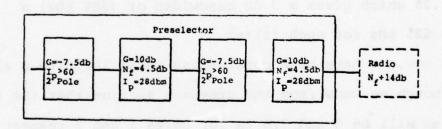
Figure 2.4   Noise Figure of Preselector and
Radio System Versus Insertion Loss
Per Filter (Case 2)

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bandwidth ratio is 1.96. Therefore, the 3 dB bandwidth will be (500 kHz) x (1.96) = 980 kHz. At 400 MHz, the worst case frequency, the loaded Q is,

$$Q_L = \frac{f_c}{BW_{3\ db}} = 408$$

where, Q_L is the loaded Q,

$BW_{3\ db}$ is the 3 dB bandwidth, and

f_c is the center frequency.

For the second case shown in Figure 2.4, using 2 pole filters we need to determine the 1.5 dB to 3 dB bandwidth ratio. Again from reference 4, for the 2 pole Butterworth filter, this ratio is 1.25 which gives a 3 dB bandwidth of (500 kHz) x (1.25) = 625 kHz for each filter.

Up to now, we have been considering ideal infinite Q elements, although we know from the previous section that the insertion loss will be 6 and 7.5 dB for cases 1 and 2 respectively. Considering these insertion losses, the unloaded to loaded Q ratio will be approximately 2 for both cases. This has no measurable effect on the single pole resonator but rounds the corner of the 2 pole filter and increases the 1.5 dB to 3 dB bandwidth ratio to 1.41. Therefore, the worst case loaded Q for the two pole case will be,

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$$Q_L = \frac{f_c}{BW_{3\ db}} = \frac{400\ MHz}{(500\ kHz)\ (1.41)} = 567.$$

D.   The Intermodulation Product Requirements

In the previous section, we calculated the bandwidth of the preselector. In this section, we will check to see if this bandwidth is narrow enough to meet the intermodulation requirements.

If two signals are applied to the preselector at frequencies 1.5 and 3 MHz above center frequency, then the non-linearities of the amplifiers will produce an interfering intermod product at center frequency. This product's level is a function of the preselector's filter bandwidth and insertion loss and the amplifier's gain and intercept point. Previous test data shows that pin diode switched filters have intercept points of approximately +80 dBm which is so much higher than the best available amplifiers ($\approx$ +27 dBm), that the filter's non-linearity can be neglected.

The intermod product level can be calculated as follows,

$$I_m = (3B - 2I_p) + 2\ (A - B)$$

where, I_m = the intermod product level at the device's output (dBm),

   A   = the power level of the strongest signal,

   B   = the power level of the weakest signal,

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I_p = the intercept point of the device at the output (dBm).

This equation is plotted in Figure 2.5. It shows that for case one using single pole filters, the 3 dB bandwidth must be less than 700 kHz in order to meet the intermod specification. In the previous section, we determined the single pole filters could be 980 kHz wide and still meet the 3 dB total preselector bandwidth goal. Since the intermod specification is the narrowest, it is the determining factor for this case.

For case two using double pole filters, Figure 2.5 shows the bandwidth must be narrower than 1.625 MHz. This is considerably wider than the bandwidth required due to the previous section. Therefore, for this case the 3 dB bandwidth is the constraining factor.

E.   Determining the Lowest Unloaded Q Requirement

We can now determine which type of preselector requires the lowest unloaded Q. In the previous sections, the insertion loss and bandwidth for case one using single pole filters must be 6 dB and 700 kHz respectively. For case two using two pole filters, the insertion loss and bandwidth must be 7.5 dB and 705 kHz respectively.

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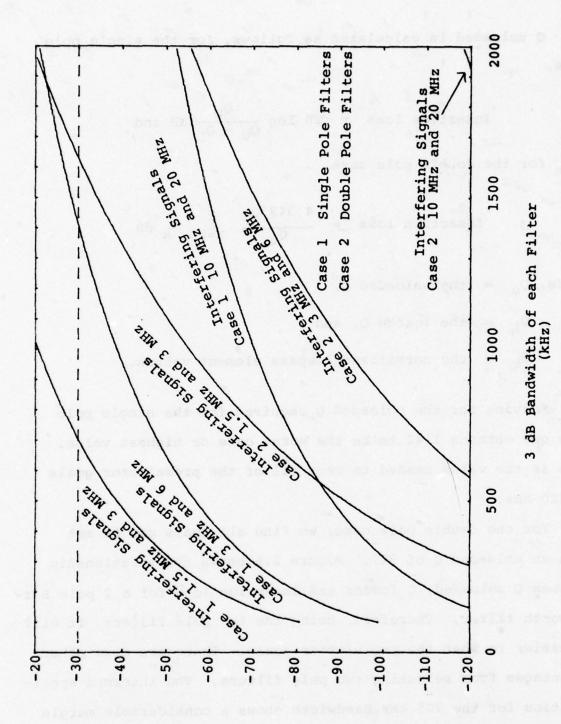
Figure 2.5 Intermodulation Product
Versus Filter Bandwidth
Prediction

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Q unloaded is calculated as follows, for the single pole case,

$$\text{Insertion loss} = 20 \log \frac{Q_u}{Q_u - Q_L} \text{ dB and}$$

for the double pole case,

$$\text{Insertion loss} = \frac{4.343 \, Q_L}{Q_u} \sum_{K=1}^{n} G_k \text{ dB}$$

where, Q_u = the unloaded Q,

Q_L = the loaded Q, and

G_k = the normalized lowpass element values.

Solving for the unloaded Q required for the single pole case one obtains 1142 to be the worst case or highest value. This is the value needed to meet all of the preselector goals at 400 MHz.

For the double pole case, we find all goals can be met with an unloaded Q of 927. Figure 2.6 shows the relationship between Q unloaded, Q loaded and insertion loss for a 2 pole Butterworth filter. Therefore, using the two pole filters, it will be easier to meet the preselector goals. There are also other advantages from selecting two pole filters. The intermod specification for the 705 kHz bandwidth shows a considerable margin and the out-of-band rejection will be substantially better than for single pole filters.

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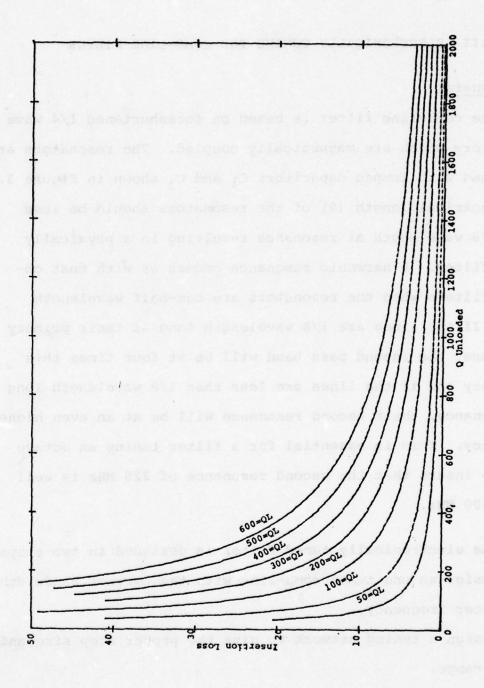
Figure 2.6 Insertion Loss Versus Q Loaded
Versus Q Unloaded for 2 Pole
Butterworth Filter

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III ELECTRONICALLY TUNING THE COMB-LINE FILTER

A. General

The comb-line filter is based on foreshortened 1/4 wave
resonators which are magnetically coupled. The resonators are
end tuned with lumped capacitors C_1 and C_j shown in Figure 3.1.
The electrical length (θ) of the resonators should be less
than 1/8 wavelength at resonance resulting in a physically
small filter. A harmonic resonance occurs as with most co-
axial filters when the resonators are one-half wavelength
long. If the lines are 1/8 wavelength long at their primary
resonance, the second pass band will be at four times this
frequency and if the lines are less than 1/8 wavelength long
at resonance, their second resonance will be at an even higher
frequency. This is essential for a filter tuning an octave
band to insure that the second resonance of 225 MHz is well
above 400 MHz.

The electronically tuned filter is designed in two steps:
1. Design an end tuned comb-line with the correct bandwidth
and center frequency.
2. Design a tuning network to give the proper step size and
tuning range.

The design equations for step one are as follows where
step two is covered in a following section.[5]

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c_1 c_j

$n=0$ θ $n=1$ $n=j$ $n=j+1$
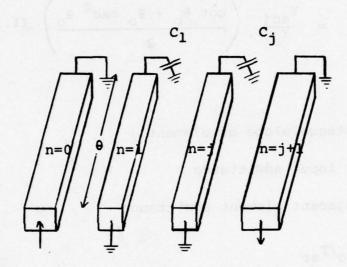
FIGURE 3.1   General Comb Line Filter

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B.   Comb-Line Filter Design

First, one must calculate the per unit length capacitances of the filter elements. The location of these is shown in Figure 3.2. Equation 1 gives the normalized susceptance slope of resonator j shown in Figure 3.1,

$$
\frac{b_j}{Y_A}\Bigg|_{j=1 \text{ to } n} = \frac{Y_{adj}}{Y_a}\left(\frac{\cot\theta_o + \theta_o \csc^2\theta_o}{2}\right) \quad (1.)
$$

where:

b_j = Susceptance slope of element j

Y_a = Filter input admittance

Y_{adj} = The adjacent element admittance

θ_o = $\pi/2\ f_o/F_{sr}$

f_o = Operating frequency and

F_{sr} = Resonant frequency of the line.

The impedance of the resonator was chosen as 77 ohms since this gives the highest unloaded Q in a coaxial line form.

Equation 1 is expanded as follows,

$$
\frac{b_1}{Y_A} = \frac{\dfrac{1}{Z\ \text{resonator}}}{\dfrac{1}{Z\ \text{input}}}\left(\frac{\cot\left(\dfrac{\pi}{2}\ \dfrac{f}{F_{sr}}\right) + \left(\dfrac{\pi}{2}\ \dfrac{f}{F_{sr}}\right)\left(\dfrac{\dfrac{1}{\sin \dfrac{\pi}{2}\ \dfrac{f}{F_{sr}}}}{}\right)^2}{2}\right)
$$

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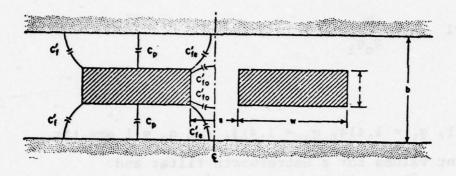
FIGURE 3.2   Capacitance and Dimension Placement
for Rectangular Elements

Where,

   b is the internal cavity height

   t is the thickness of the resonator

   s is the spacing between elements

   w is the width of resonator

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For a two pole filter,

$$\frac{b_1}{Y_A} = \frac{b_2}{Y_a} \qquad \text{and}$$

the input conductance is

$$G_{T1} = \frac{\omega b_1}{g_o g_1} \quad ,$$

where

$g_o = 1$, $g_1 = 1.414$, $g_2 = 1.414$, and $g_3 = 1$ are the element values for a Butterworth filter and

ω is the 3 dB bandwidth divided by the center frequency.

The value of the admittance inverter is

$$J_{j,\ j+1} \bigg|_{j=1} = \omega \left[\frac{b_j}{g_j} \frac{b_{j+1}}{g_{j+1}} \right]^{1/2}$$

We can now calculate the capacitance per unit length of the input transformer,

$$\frac{C_o}{\epsilon} = \frac{376.7\ Y_A}{\sqrt{\epsilon_r}} \left[1 - \left(\frac{G_{T1}}{Ya} \right) \right]^{1/2}$$

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and the capacitance per unit length of the next resonator,

$$\frac{C_1}{\varepsilon} = \frac{376.7}{\sqrt{\varepsilon_r}} \left[Y_{al} - Y_A + G_{T1} - J_{12} \tan \theta_o \right] + \frac{C_0}{\varepsilon}$$

where:

ε is the absolute dielectric constant of free space and

$\varepsilon_r = 1.0$ when the dielectric is air.

The coupling capacitance per unit length between elements

zero and one is,

$$\frac{C_{01}}{\varepsilon} = \frac{376.6 \, Y_A}{\sqrt{\varepsilon_r}} - \frac{C_0}{\varepsilon}$$

and between elements one and two is

$$\frac{C_{12}}{\varepsilon} = 376.7 \, J_{12} \tan \theta_o$$

C. Determining the Dimensions

The normalized distance between elements zero and one is

$\frac{S_{01}}{b}$ and is found from Figure 3.3 using the calculated value

of $\frac{C_{01}}{\varepsilon}$ and t/b which is the thickness to height ratio. The

normalized distance between the resonantors, element one and

two is $\frac{S_{12}}{b}$ from Figure 3.3 using the calculated value of

$\frac{C_{12}}{\varepsilon}$.

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The fringing capacitance $\dfrac{C'_{fe}}{\varepsilon}$ on the outside edge of the

transformer element zero is found from Figure 3.4. The even

mode finging capacitances between elements are found from

Figure 3.3 using $\dfrac{S_{01}}{b}$ and $\dfrac{S_{12}}{b}$.

With these fringing capacitances we can solve for the norma-

lized element widths,

$$\frac{\omega_0}{b} = \frac{1}{2}\left(1 - \frac{t}{b}\right)\left[\frac{1}{2}\frac{C_0}{\varepsilon} - \frac{C'_{fe_{01}}}{\varepsilon} - \frac{C'_{fo}}{\varepsilon}\right] \text{ for element zero}$$

and

$$\frac{\omega_1}{b} = \frac{1}{2}\left(1 - \frac{t}{b}\right)\left[\frac{1}{2}\frac{C_1}{\varepsilon} - \frac{C'_{fe_{01}}}{\varepsilon} - \frac{C'_{fe_{12}}}{\varepsilon}\right] \text{ for element one.}$$

The t/b ratio can be chosen somewhate arbitrarily although

for this filter it was selected to give optimized distances
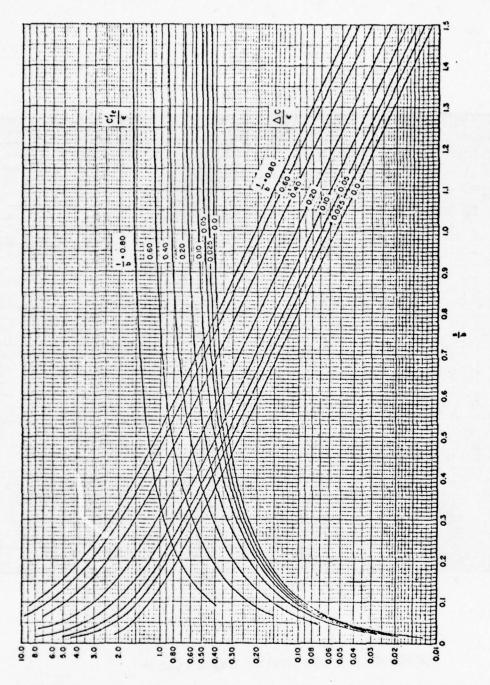
for parts placement and assembly.

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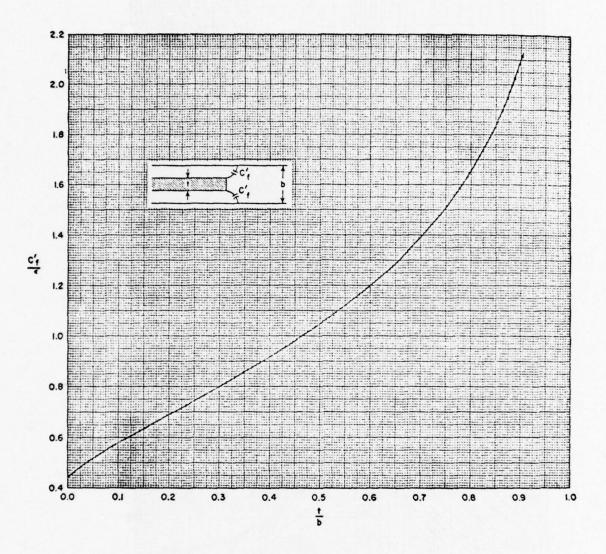FIGURE 3.3 Fringing and Interbar Capacitances
for Coupled Rectangular Bars

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FIGURE 3.4 Fringing Capacitance

IV   THE ELECTRONIC TUNING NETWORK

A.   General

The basic concept of the tuning network is that high Q
capacitors can be switched in binary steps with low loss PIN
diodes to provide the necessary capacitance step size and tuning
range. After due consideration, it was decided to tap all capa-
citors at the same point on the resonator so that spurious res-
onances would be eliminated and so that even tuning increments
would be easier to achieve. Naturally, the alternative is to use
the same value of capacitance spaced at varying distances along
the resonator to give binary tuning steps. The latter technique
was used on an earlier contract.[1, 2] Coupling between tuning
elements was reported to have caused the binary step size
to vary versus frequency. It was felt that with our much smaller
packaging volume this coupling between capacitors would be sub-
stantially increased causing unmanageable tuning difficulties.

B.   Determining the Capacitance Tuning Range

The maximum tuning capacitance, although a function of
many variables, is ultimately determined by the tuning step size
and frequency range. The procedure for calculation of the mini-
mum and maximum capacitance range is as follows:

1. First choose a PIN diode with high Q off capacitance
and low on resistance.

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2.   The minimum tuning step (250 kHz) is used to determine
the number of binary tuning bits.

3.   The total stray capacitance is calculated from the off
capacitance of each pin diode and the number of tuning bits.

4.   The self resonant frequency of the resonator is calcula-
ted based on the maximum tuning frequency and stray capacitance
of the tuning network.

C.   Pin Diode Selection

A large number of different pin diodes were considered for
use in the tuning network. Figures 4.1, 4.2 and 4.3 show the Q
variation for the combination of a pin diode in series with a
tuning capacitor versus the value of the tuning capacitor which
the diode switches. The curves illustrate the predicted Q for
both the forward biased and reverse biased states of the pin
diode.
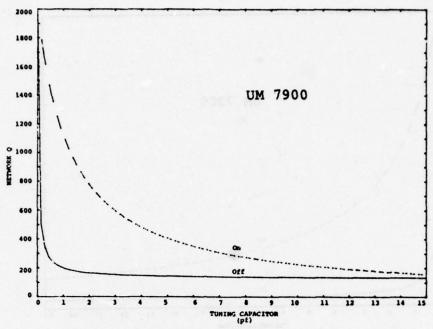
1.   The Forward Biased Q

In the forward biased state, the curve consists of
the following,
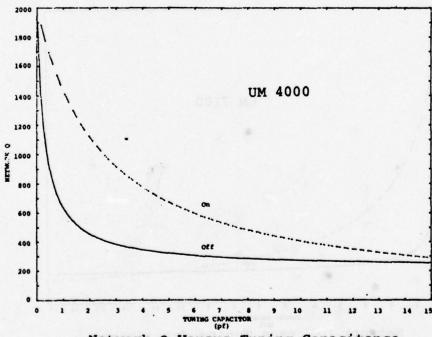
$$Q_{eff} = \frac{Q_t}{1 + 2\pi f_c C_t R_d Q_t}$$

where,

f_c = the operating frequency,

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UM 7900

Network Q Versus Tuning Capacitance

UM 4000

Network Q Versus Tuning Capacitance

FIGURE 4.1

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UM 7200

Network Q Versus Tuning Capacitance

UM 7100

Network Q Versus Tuning Capacitance
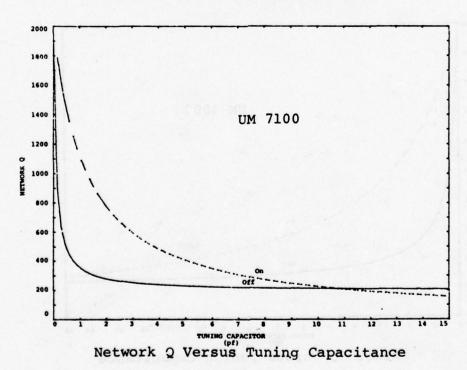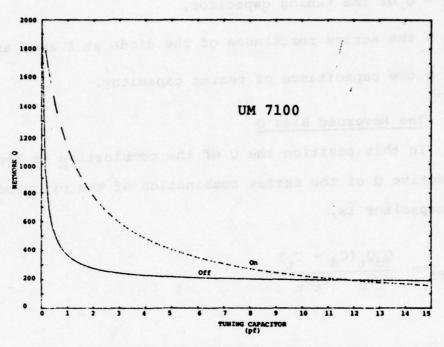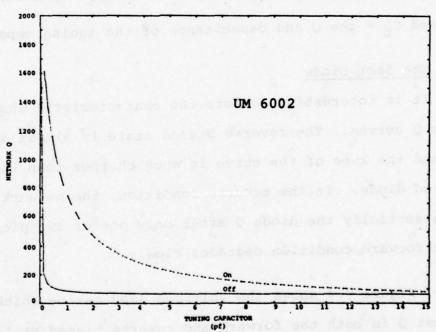
FIGURE 4.2

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Network Q Versus Tuning Capacitance

Network Q Versus Tuning Capacitance

FIGURE 4.3

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Q_t = Q of the tuning capacitor,

R_d = the series resistance of the diode at 1 amp, and

C_t = the capacitance of tuning capacitor.

2. The Reversed Bias Q

In this position the Q of the combination is lower. The effective Q of the series combination of the pin diode and tuning capacitor is,

$$Q_{eff} = \frac{Q_d Q_t (C_d + C_t)}{C_d Q_d + C_t Q_t}$$

where,
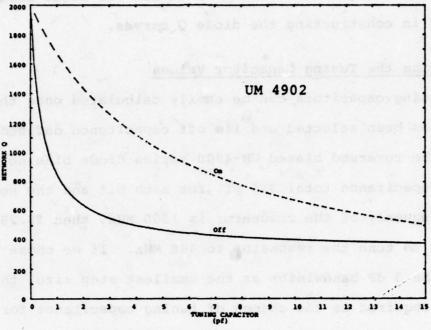
Q_d and C_d = the Q and capacitance of the pin and diode and

Q_t and C_t = the Q and capacitance of the tuning capacitor.
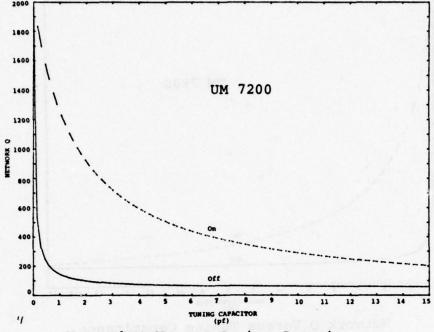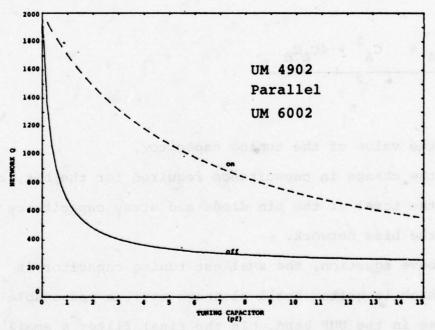
3. The Best Diode

It is interesting to note the characteristic shape of the diode Q curves. The reverse biased state is almost always lower Q and the knee of the curve is much sharper than the forward biased diode. In the reverse condition, the network Q reaches essentially the diode Q after only one or two picofarads while the forward condition degrades slowly.

The top of Figure 4.4 shows the Unitrode 4900 series which has the highest Q in both the forward and reverse biased states; therefore, it was selected as the pin diode switch for the

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Network Q Versus Tuning Capacitance

UM 4902

(axes: NETWORK Q vs TUNING CAPACITOR (pf), curves labeled "on" and "off")

Network Q Versus Tuning Capacitance

UM 4902
Parallel
UM 6002

(axes: NETWORK Q vs TUNING CAPACITOR (pf), curves labeled "on" and "off")

FIGURE 4.4

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tuning network. Figure 4.5 lists the on and off diode resistance values used in constructing the diode Q curves.

D. Determing the Tuning Capacitor Values

The tuning capacitors can be easily calculated once the pin diode has been selected and its off capacitance defined. If we assume the reversed biased UM-4900 series diode bias network and stray capacitance total 3.5 pf. for each bit and the self resonant frequency of the resonator is 1200 MHz, then 18.26 pf is required to tune the resonator to 400 MHz. If we chose 250 kHz (half the 3 dB bandwidth) as the smallest step size, then 0.02 pf is required as the change of tuning capacitance for this bit. The value of the tuning capacitor is calculated as follows,

$$C_t = \frac{C_\Delta + \sqrt{C_\Delta^2 + 4C_\Delta C_d}}{2}$$

where,

C_t is the value of the tuning capacitor,

C_Δ is the change in capacitance required for the bit, and

C_d is the total of the pin diode and stray capacitance of the bias network.

Using the above equation, the smallest tuning capacitor is 0.2748 pf which is quite small although still a reasonable value for use in the UHF band. In the final filter a small variable capacitor was used which was designed for us by Johanson

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TUNING NETWORK ELEMENT VALUES
FOR TUNING NETWORK Q

DIODE	DIODE OFF CAPACITANCE	DIODE OFF PARALLEL RESISTANCE	DIODE ON SERIES RESISTANCE	TUNING CAPACITOR Q
UM4902	2.8pf	60K ohm	.05 ohm	2000
UM7900	.6pf	110K ohm	.20 ohm	2000
UM4000	2.8pf	40K ohm	.10 ohm	2000
UM7200	2.0pf	13K ohm	.15 ohm	2000
UM7100	1.0pf	100K ohm	.20 ohm	2000
UM6002	.4pf	80K ohm	.40 ohm	2000
UM4302	1.3pf	70K ohm	.20 ohm	2000
UM4902 in parallel with UM6002	3.2pf	35K ohm	.044 ohm	2000

FIGURE 4.5

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to tune from 0.1 to 0.5 pf. This capacitor was simply tuned to give the correct minimum step size.

The larger value tuning capacitors are determined using the same equation by using binary delta capacitance steps. These are given in Figure 4.6 for the 12 tuning bits.

E.   The Bias Networks

A proportionately large amount of time was spent on the design and analysis of the bias networks. They are probably the most critical element in the design of the PIN diode switched filter. They effect frequency step size, frequency tracking, temperature stability and insertion loss, in other words, all major specifications of the filter.

The two major bias network alternatives were both computer modeled and breadboard tested.

1.   Bias Chokes

This technique uses a series   RF choke in the DC control line which presents a high impedance to the tuning net-work. A schematic of RF choke biasing is shown in Figure 4.7. The computer analysis indicates that RF choke inductance values greater than 1 uH are necessary in order to not distort the fil-ter's tuning curve. Ideally about 10 uH gives negligible tun-ing distortion although 1 uH chokes can be used if the capaci-tances of the larger tuning bits are reduced. This large

TUNING NETWORK CAPACITOR VALUES

Bit	Delta Capacitance (pf)	Diode Capacitance (pf)	Tuning Capacitance (pf)
0	.02	3.5	.2748
1	.04	3.5	.3947
2	.08	3.5	.5707
3	.16	3.5	.8326
4	.32	3.5	1.2303
5	.64	3.5	1.8505
6	1.28	3.5	2.8512
7	2.56	3.5	4.5355
8	5.12	3.5	7.5071
9	10.24	3.5	12.9975
10	12.1622	3.5	15.0000
11	12.1622	3.5	15.0000

Total Delta Capacitance = 44.784 pf

Maximum Capacitance = 63.045 pf

Minimum Capacitance = 18.261 pf

FIGURE 4.6

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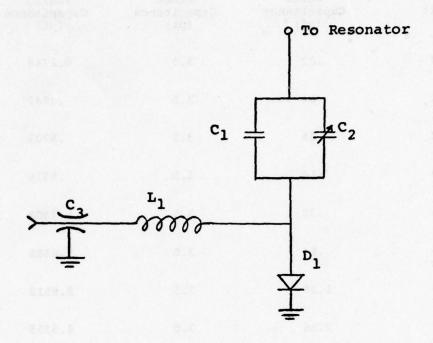To Resonator

C_1 C_2

C_3 L_1

D_1

Figure 4.7 Tuning Network
with Bias Choke

inductance value is required because the RF choke changes
impedance versus frequency, and its impedance must be substan-
tially greater then that of the tuning capacitor. Since this
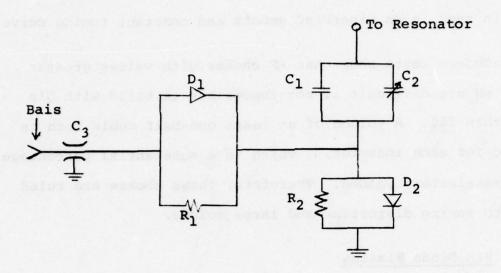contract requires the filter to tune almost an octave, this

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impedance change is much greater than if the filter tuned a narrow band. The impedance change of the RF choke causes tuning bits to change in effective capacitance versus frequency and results in gaps in an otherwise smooth and constant tuning curve.

Breadboard tests show that RF chokes with values greater than 200 nH are difficult if not impossible to build with Q's greater than 200. A volume of at least one-half cubic inch is allocated for each inductor, which is a substantial percentage of the preselectors volume. Therefore, these chokes are ruled out due to tuning distortion and large volume.

2. Pin Diode Biasing

This technique uses a small PIN diode in series with the DC control line as a switch to bias the larger PIN. The schematic of this network is shown in Figure 4.8.

In the forward biased direction, the small PIN diode (a UM-6002) has 0.4 ohms series resistance which lowers the total series resistance to ground by about 10 percent and improves the forward biased Q of the network. In the reverse biased direction, the small diode appears as an 0.4 pf capacitor in parallel with a resistance (Rp) of 80K ohms. The advantage here is that the network has no inductors which can distort the tuning curve with spurious resonances, although the reverse biased Q of the small PIN diode is not as good as the Q of the

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To Resonator

Bais

C_1 Ceramic tuning capacitor

C_2 Johson Giga Trim used to fine tune C_{Tune} bit

C_3 1000pf feed thru capacitor

D_1 Bais pin diode UM6002

D_2 Switch pin diode UM4902

R_1, R_2 Bais divider resistors 3.9megohm

Figure 4.8   Tuning Network
with Bias Diode

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large diode and lowers the reversed bias Q of the network.
The bottom of Figure 4.4 shows a plot of the network Q using
the large and small pin diodes in parallel versus the value of
tuning capacitance.

V THE COMPUTER MODEL AND PERFORMANCE PREDICTION

A. General

The electronically tuned filter was computer modeled to optimize its performance versus frequency. Initially component manufacturer's test data was used for pin diode and capacitor modeling, although as the contract progressed, the models were refined as more of our test data became available. Overall, the computer model was invaluable for finding and eliminating spurious resonances in the bias networks and optimizing and maintaining the insertion loss flat versus frequency.

B. Filter Modeling

The general comb line filter structure is shown in Figure 5.1. This network can be easily modeled with parallel transmission line elements as shown in Figure 5.2. Although by transforming this to an open wire model, shown in Figure 5.3, computer running time is reduced since fewer transmission line elements are used. The open wire model can also be transformed to an RLCX network at a single frequency. This model, as with the others, assumes that the non-adjacent elements (such as 1 and 3) have a value of coupling so small that it can be ignored. This assumption is valid for the bandwidths required for this contract.
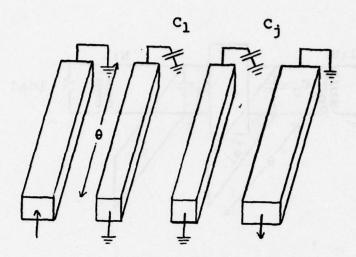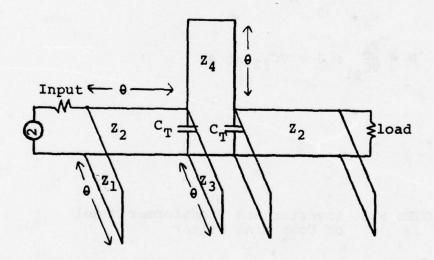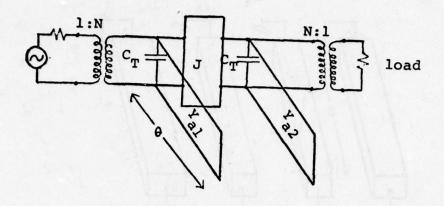
FIGURE 5.1 General Comb Line Filter

$$z_1 = \frac{1}{VC_\emptyset} \; ; \; z_2 = \frac{1}{VC_{\emptyset 1}} \; ; \; z_3 = \frac{1}{VC_1} \; ; \; z_4 = \frac{1}{VC_{12}}$$

FIGURE 5.2 Open Wire Model of
Comb Line Filter

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$$Y_{a1} = Y_A \left[1 - \left(\frac{VC_{\phi 1}}{Y_A} \right)^2 \right] + V(C_1 + C_{12} - C_{\phi});$$

$$N = \frac{Y_A}{VC_{\phi 1}}; \quad J = VC_{12} \text{ at } \theta$$

FIGURE 5.3   Inverter and Transformer Model
of Comb Line Filter

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The next design step involves selecting the correct tap
point on the resonator for the tuning network. As shown in
Figure 5.4, the resonators are tapped at approximately their
center. As the tap point is moved closer toward the grounded
end of the resonator, the driving impedance of the tuning net-
work is lowered. This increases the insertion loss at the
lower end of the tuning range and decreases the insertion loss
at the higher end of the tuning range. Therefore, the tap
point can be selected at a point to give flat insertion loss
versus frequency. To properly analyze the resonators, they
are considered disjointed at the tap point and considered as
two cases. Figure 5.5 shows the resonators from the shorted
end to the tap point of the tuning network. Figure 5.6
considers the resonators from the tap point to the open end.

The transmission lines can now be simplified at an indi-
vidual frequency and converted to their LC model equivalent
circuits. The model of the tuning network, including the
12 capacitor tuning bits and bias networks, is added and the
overall response is obtained from a network analysis computer
program.

Figure 5.7 shows the final model of the filter. The
shorter input and output transformers are grounded in the
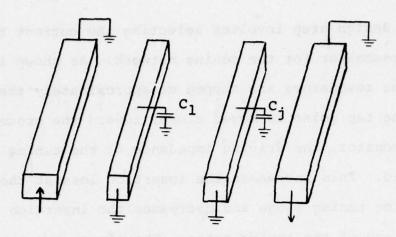vicinity of the tap point. Constant coupling versus

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FIGURE 5.4   Comb Line Filter with
Center Tapped Resonators
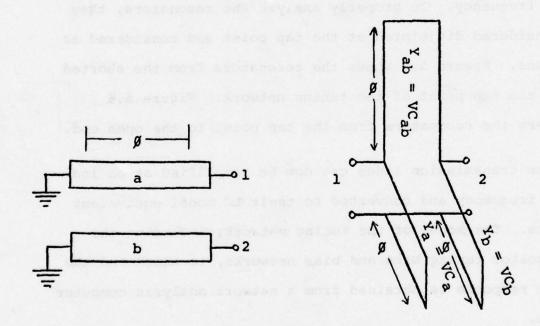
FIGURE 5.5   Transmission Line Pair
with Shorted Ends

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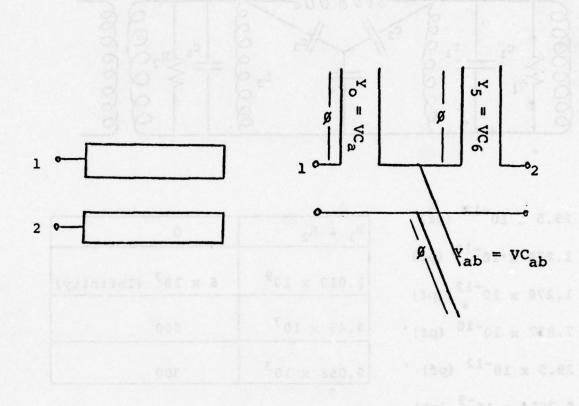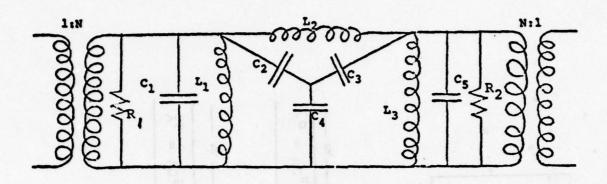
FIGURE 5.6   Transmission Line Pair
with Open Ends

C_1	29.5×10^{-12} (pf)
C_2	1.278×10^{-12} (pf)
C_3	1.278×10^{-12} (pf)
C_4	7.837×10^{-10} (pf)
C_5	29.5×10^{-12} (pf)
L_1	8.7614×10^{-9} (mh)
L_2	5.439×10^{-6} (mh)
L_3	8.7614×10^{-9} (mh)

R_1 & R_2	Q
1.012×10^9	6×10^7 (Infinity)
8.43×10^3	500
5.058×10^3	300

FIGURE 5.7   Inductor Capacitor and Resistor
Model Comb Line Filter

frequency results since changes in currents at the tap point
are located near the grounded end of transformer.

C.   Filter Performance Prediction

The results of the computer model are plotted in Figure
5.8. This graph illustrates how insertion loss and bandwidth
change versus unloaded Q. As to be expected, the 3 dB band-
width widens and the insertion loss increases as the unloaded
Q is reduced.

A special computer program developed on an earlier project
tested the computer models automatically in each pin diode
state. Figure 5.9 plots unloaded Q versus frequency using
pin diode and capacitor models developed from test data pro-
vided by the component manufacturers. These parameters are
shown in Figure 5.10.

Later in the contract, these parts were measured in the
filter. The measured Q's were somewhat lower than anticipated
due to the earlier data. The parallel off resistance (R_p) of
the pin diodes was measured in the filter network to be about
30KΩ instead of the 60KΩ value that was expected. Figure
5.11 shows the unloaded Q with the same parameters as in
Figure 5.10, except R_p is 30KΩ . This degraded the performance
at the high end of the UHF band and was responsible for the
filter not meeting the bandwidth and insertion loss goals.

Note that at 400 MHz all of the tuning diodes are turned off and the diode R_p is by far the major contributor to insertion loss.

Insertion Loss
(dB)

Qunl = 6 x 10^7
(Infinity)

Qunl = 500

Qunl = 300

Frequency
(MHz)

319.0   319.5   320.0   320.5   321.0

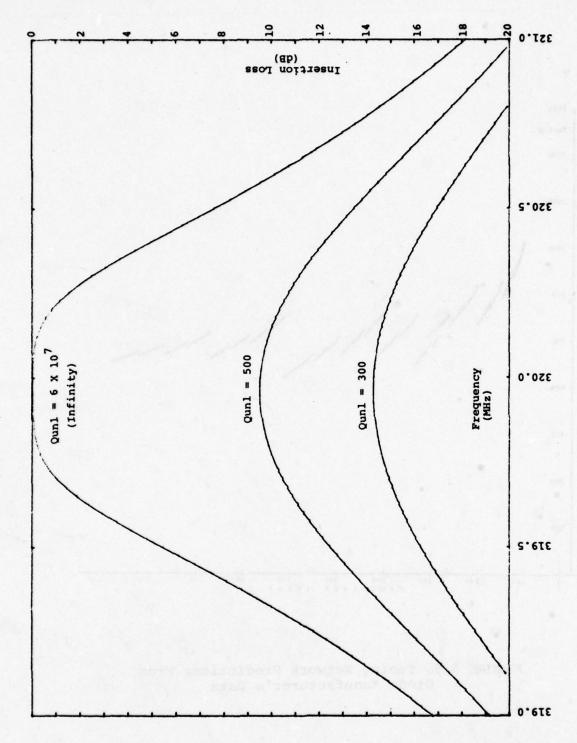0   2   4   6   8   10   12   14   16   18   20
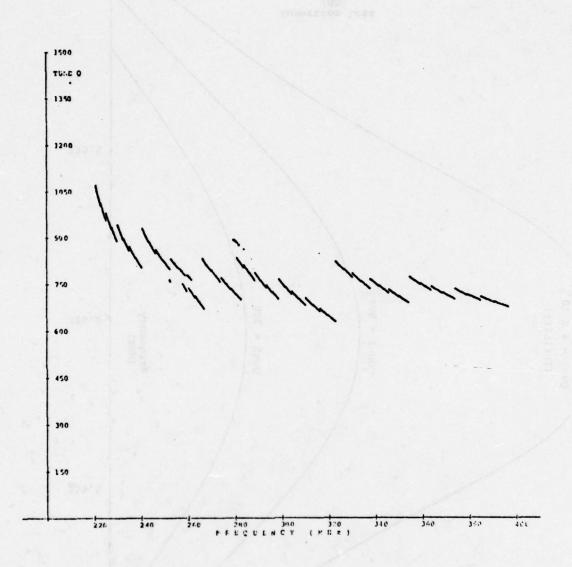
FIGURE 5.8   Insertion Loss Versus Capacitor Q

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FIGURE 5.9 Tuning Network Predictions From
Diode Manufacturer's Data

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Parameter	Bit 11	Bit 10	Bit 9	Bit 8	Bit 7	Bit 6	Bit 5	Bit 4	Bit 3	Bit 2	Bit 1	Bit 0
Tuning Capacitor	15.000	15.000	12.997	7.507	4.535	2.851	1.850	1.230	.832	.570	.395	.275
Capacitor R_s (Ω)	.018	.018	.021	.034	.056	.089	.130	.207	.306	.447	.649	.928
Diode R_s (on)	.036	.036	.036	.036	.036	.036	.036	.036	.036	.036	.036	.036
Diode R_p (off)	60 K	60 K	60 K	60 K	60 K	60 K	60 K	60 K	60 K	60 K	60 K	60 K
Diode C_p (off)	3.50	3.50	3.50	3.50	3.50	3.50	3.50	3.50	3.50	3.50	3.50	3.50

FIGURE 5.10 Tuning Capacitors and Pin Diode
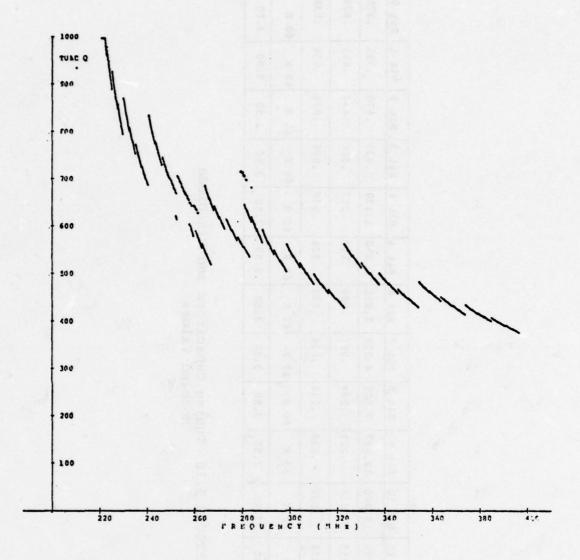Modeled Values

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Figure 5.11 Tuning Network Prediction
from Measured Data

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VI PRESELECTOR FILTER DIMENSIONS

A. General

This section discusses the physical dimensions selected
for the comb line filters used in the UHF preselector. The
size of the present tuning networks, the length of the
resonators, and the element-to-element spacing determine
the minimum inside dimensions. Choosing bias diodes over
bias chokes as discussed in Section IV considerably reduces
the overall outside dimensions of the preselector. The
chokes were previously placed on the outside of the filter
while the bias diodes are small enough to be placed on the
inside.

B. The Minimum Filter Dimensions

The minimum inside dimension of the comb line filter is
determined by the size of components mounted on the interior
wall. Spacing between the resonator and the side wall must
be at least 0.3 inch to allow adequate room for the pin
diode and capacitor network. The filters are greater than
2.34 inches long to accommodate the length of the resonators.
Although not a problem for this contract, the resonators must
be greater than 0.16 inches wide to allow threads for ease
of assembly. The diameter of the tuning network should be

substantially less than the resonators length to prevent
interference with the cavity fields.

C.   The Large Filter's Construction

During this contract two filters were designed and
tested. The larger filter was used in the input section of
the preselector where the electrical requirements are more
critical. This filter's dimensions are shown in Figures
6.1, 6.2, 6.3 and 6.4. The size was chosen for high Q and
ease of construction. The pictures in Figure 6.5 show the
inside of the filter with conventional transformers. Later in
the contract, shorter input transformers shown in Figure 6.4
were tested and since they had improved input impedance
versus frequency were selected.

D.   The Tuning Networks

One of the tuning networks is shown in Figure 6.6. Each
network incorporates six tuning bits which are mounted on a
plate that can be removed from the side wall. The plates have
a fairly large diameter to maintain a low RF resistance with
the side wall. Two plates are used on each side of the
resonator to give a total of 12 tuning bits. A post with a
thin disk is attached to the resonator and tuning capacitors
are soldered to the disk. The post is attached to the center
of the resonator. The disk is constructed of several skin
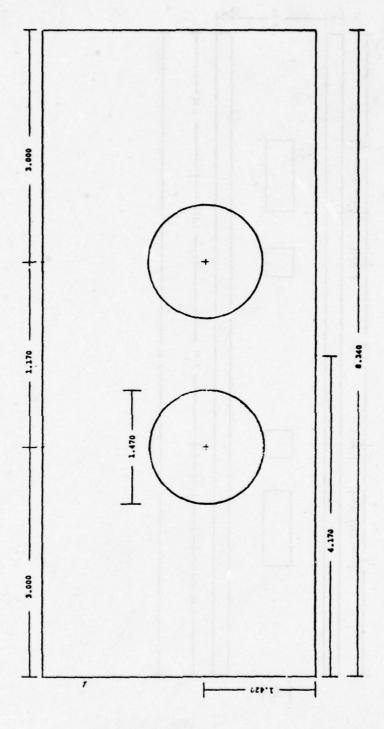
58

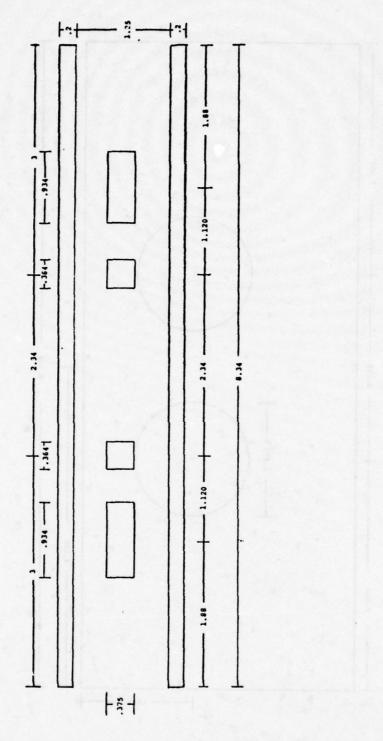Figure 6.1   Top Plate of Large
Comb Line Filter

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Figure 6.2   End Dimensions of Large
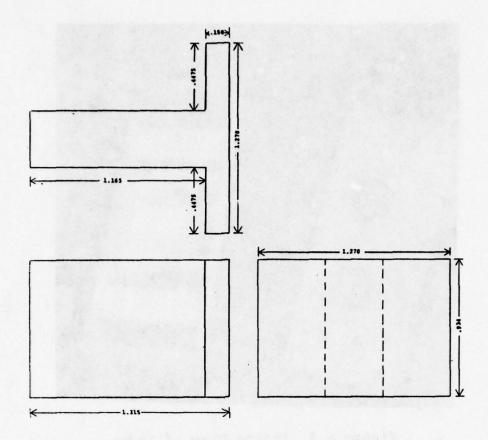Comb Line Filter

60

Figure 6.3　Reduce Length Transformer
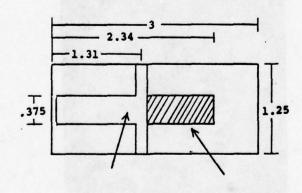for Large Filter

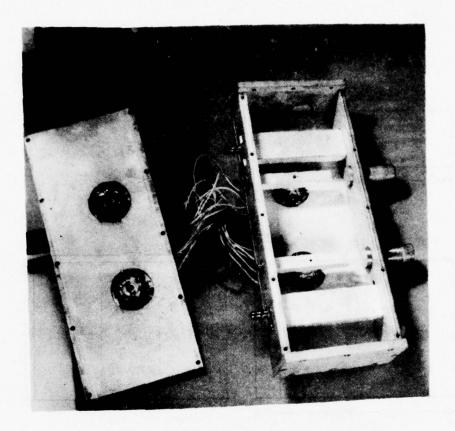Figure 6.4　Side View of Short
Transformer in Large
Filter

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Figure 6.5   Inside View of Large
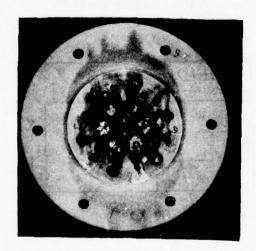   Comb Line Filter

Figure 6.6   Tuning Plate for Large
             Comb Line Filter

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depths of silver plated on copper foil to provide stress relief for the tuning network because the diodes have a tendency to crack under a very low axial shear stress. Figure 6.7 shows the method used to connect the tuning network to the resonator.

E.   The Small Filter's Construction

The small filter was designed with the restriction that the largest dimension be less than five inches. Figures 6.8, 6.9 and 6.10 illustrate the construction techniques of this filter. Figure 6.11 is a drawing showing the resonator's construction. The base of the resonators are flared to provide a lower resistance RF current path. Recessing the end of the resonator assures greater contact pressure and a lower resistance RF current path. Figures 6.12 and 6.13 show the interior parts placement for this filter.

F.   The Final Preselector Package

Figure 6.14 shows the final package. Both the large filter and small filter are mounted in a chassis along with the two high dynamic range amplifiers.

Connecting Post
Connecting Disk

← **Resonator**

Tuning Capacitor
Pin Diode
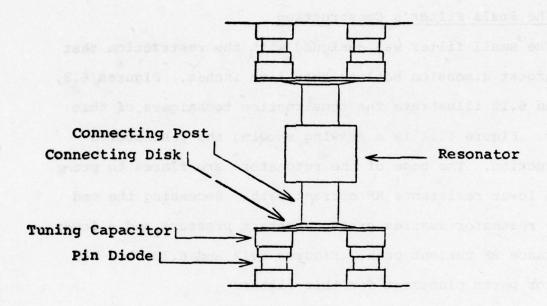
Figure 6.7   Resonator and Tuning
Network Connection
Configuration

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Figure 6.8   Outside Dimensions of Top
             Plate for Small Comb Line
             Filter

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Figure 6.9 Inside of Top Plate for
Small Comb Line Filter

Figure 6.10   End Dimensions for Small
Comb Line Filter

Figure 6.11   Resonator for Small
Comb Line Filter

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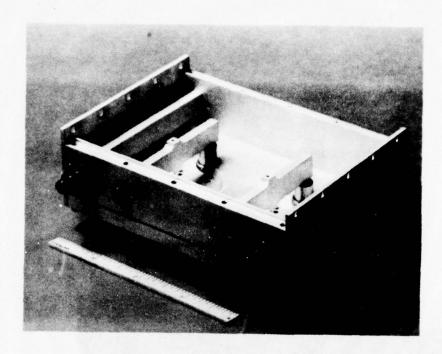
Figure 6.12   Small Comb Line Filter
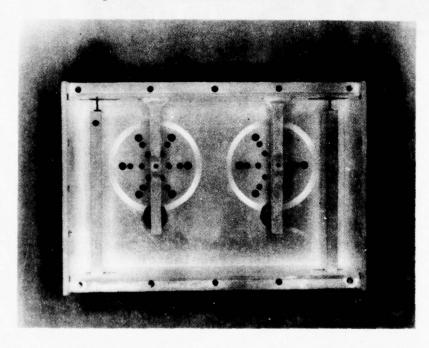
Figure 6.13   Inside View of Small
Comb Line Filter

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Figure 6.14   Electronically Tuned
UHF Preselector

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VII   PRESELECTOR MEASUREMENTS

A.   General

Due to the large number of data points required to accur-
ately characterize the preselector from 225 to 400 MHz, a
computer program was written to automatically control the
preselector and the HP-8507 network analyzer. This program
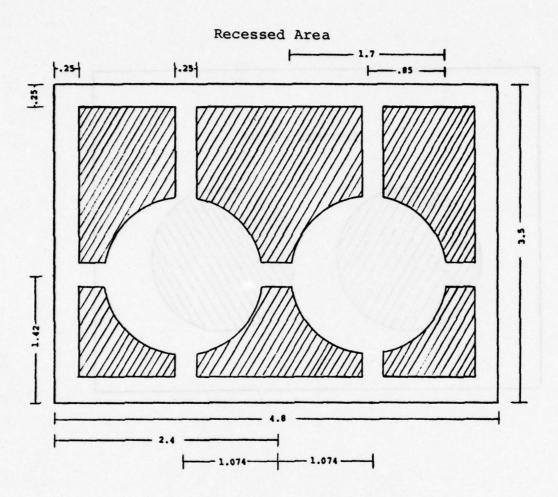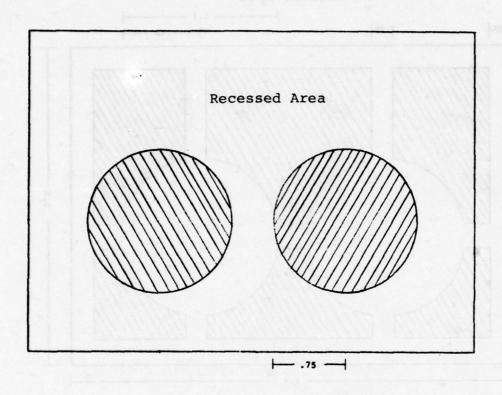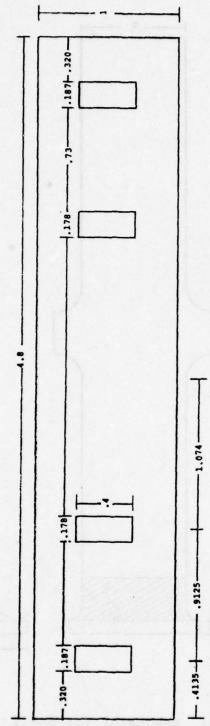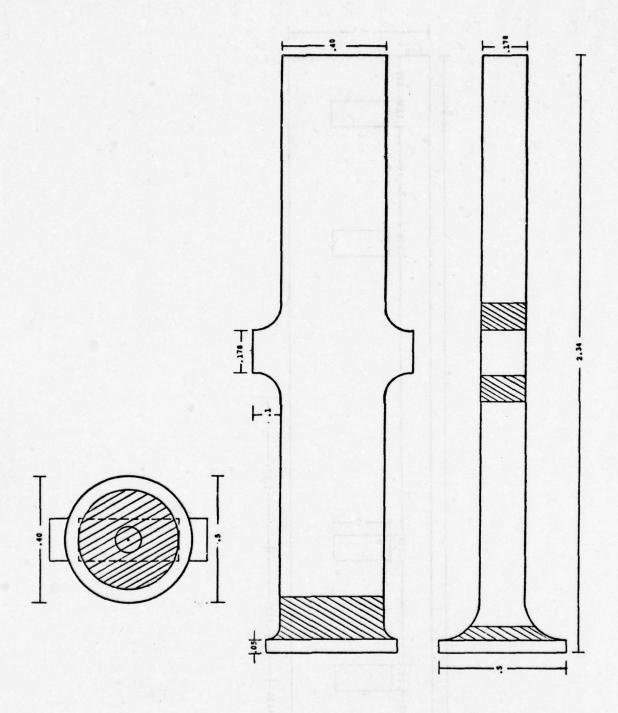was designed to plot the preselector's performance at every 10
MHz across the UHF band. This test set not only reduced the
measurement time but also gave more accurate results. Insertion
loss was measured to within 0.1 dB and the bandwidth was
characterized to within a few kilohertz.

B.   Frequency Range

The frequency tuning range was met as shown in Figure 7.1,
although the preselector's performance was better at the center
than at either the low or high end of tuning range.

C.   Bandwidth and Insertion Loss Data

Insertion loss versus frequency was measured automatically
by the HP-8507 Network Analyzer shown in Figure 7.2. One
hundred insertion loss points over ±10 MHz centered at each of
the 18 evenly spaced center frequencies were taken and stored on
magnetic tape. These are shown plotted in Figure 7.1.

The smallest tuning increment at 400 MHz is 250 kHz, while
at 225 MHz it is approximately 35 kHz.
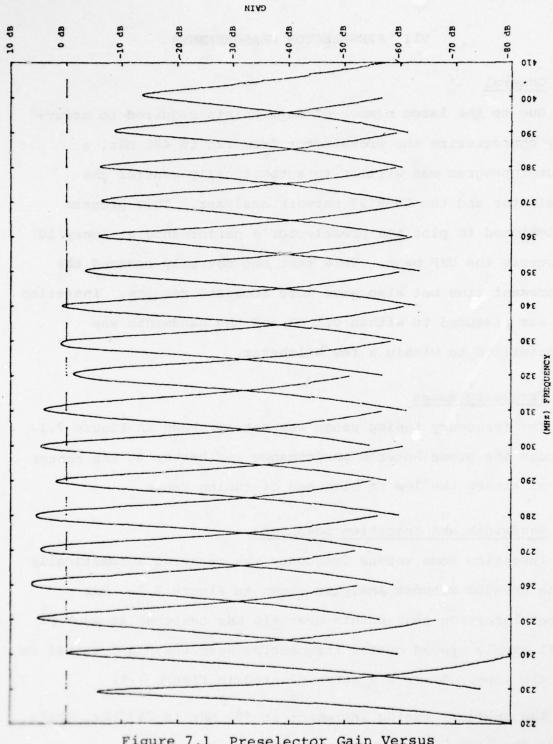
Figure 7.1 Preselector Gain Versus
Frequency from Each
Channel's Data

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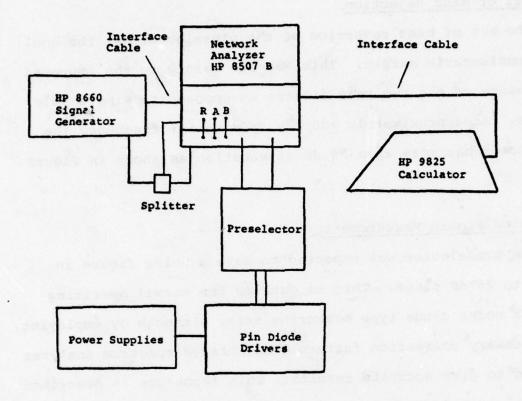
Figure 7.2   UHF Preselector Gain and
            Return Loss Measurement
            Test Setup

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Figure 7.3 is a summary table showing the return loss, gain, 3 dB, 20 dB, and 40 dB bandwidths.

D. Out of Band Rejection

The out of band rejection of the preselector met the goal by a considerable margin. This was due mainly to the improved performance of the two pole filters as opposed to single pole filters. At approximately ±20 MHz from center frequency the preselector has more than 85 dB attenuation as shown in Figure 7.4.
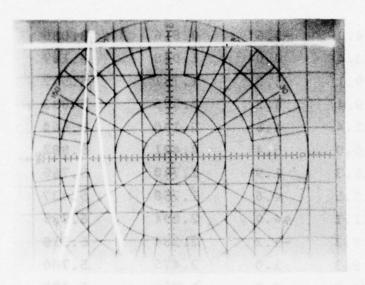
E. Noise Figure Measurement

The preselector was expected to have a noise figure in the 10 to 25 dB range. This is outside the normal operating range of noise diode type measuring sets, although by employing the necessary correction factors a calibrated spectrum analyzer was used to give accurate results. This technique is described in detail in reference 7.

The test equipment configuration is shown in Figure 7.5. The two low noise 10 dB gain amplifiers are used before the spectrum analyzer to provide a low input noise figure. Noise figure is calculated by first measuring the preselector's gain and noise power output, with the input terminated in a low noise 50Ω termination.

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Channel Freq. (MHz)	Return Loss (dB)	Gain (dB)	3 dB Bandwidth (MHz)	20 dB Bandwidth (MHz)	40 dB Bandwidth (MHz)
230	-14.0	-5.6	0.816	3.036	6.651
240	-23.4	0.0	1.048	4.107	8.040
250	- 9.1	5.2	1.781	5.572	10.911
260	- 9.4	6.2	2.525	5.893	10.995
270	-12.4	4.6	2.190	5.664	10.414
280	-10.0	5.4	2.421	5.882	10.568
290	-16.3	1.9	1.740	5.766	10.527
300	-15.5	4.6	1.758	5.437	10.531
310	-11.1	4.0	2.094	6.789	12.436
320	- 8.3	-1.2	2.384	6.318	11.635
330	- 8.3	1.0	2.475	5.780	10.926
340	-20.0	0.8	1.758	5.408	10.589
350	- 9.2	-3.4	1.778	5.800	11.027
360	- 9.2	-5.8	1.403	5.602	11.492
370	- 8.5	-6.4	1.664	6.067	12.887
380	-10.9	-6.1	1.488	5.794	12.974
390	-10.7	-8.7	1.721	6.449	14.235
400	- 5.8	-13.7	1.821	7.390	16.593

Figure 7.3   Preselector Gain and
Bandwidth Data

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Reference Line 0 dB Gain
Vertical 10 dB/Div
Horizontal Sweep 220 MHz to 410 MHz
Manual Control
Input Signal Level -33 dBm

Figure 7.4   Out-of-Band Rejection
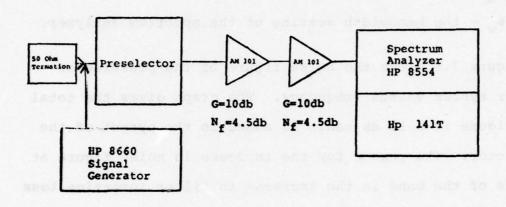   UHF Preselector Channel
             270 MHz

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Figure 7.5 Preselector Noise Figure
Measurement Test Setup

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N.F.=10 log N_O-10 log G_a-(-174 dBm/Hz - 1.7 dB* +10 log B_w)

  *The 1.7 dB is a correction factor for the noise bandwidth,
   the detector characteristics and the log scale of the spectrum
   analyzer.

where, N_O = the measured noise power,

   G_a = the preselector's gain, and

   B_w = the bandwidth setting of the spectrum analyzer.
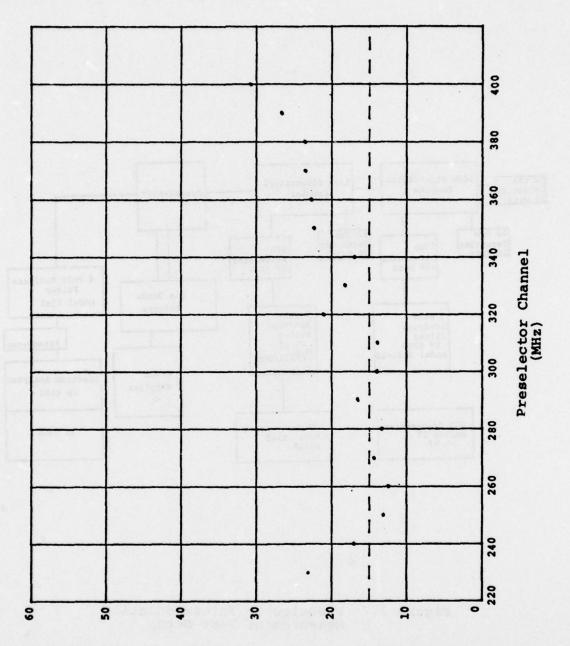
Figure 7.6 shows the noise figure of the preselector
receiver system versus frequency. The graph gives the total
noise figure if a 14 dB radio is added to the output of the
preselector. The reason for the increase in noise figure at
the ends of the band is the increase in filter insertion loss
at the ends of the UHF band.

F.   Maximum Out-of-band Signal Level

The preselector operates with +20 dBm out of band signal
levels with no problem. Additionally the 1 dB in band compress-
ion point is +17 dBm at the output. This high power capability
allows the preselector to handle even +20 dBm in band levels
without failure.

G.   Intermodulation Product Data

Extreme care must be taken in developing any test equipment
configuration for intermod tests. The incercept point of the
test setup must be higher than the intercept point of the device

Preselector plus (14db Noise Figure) Radio Noise Figure
(db)

Figure 7.6   Preselector Plus Radio
Noise Figure

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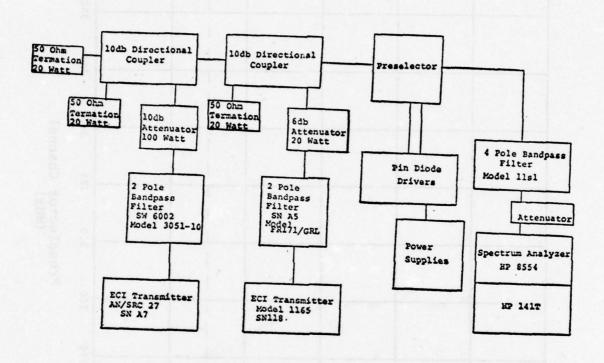
Figure 7.7   Preselector Intermodulation
Measurement Test Setup

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to be tested. The final test equipment configuration is shown in Figure 7.7.

The two +20 dBm tones are generated by UHF transmitters followed by bandpass filters and pad attenuators. The two 10 dB directional couplers are used to give additional isolation. These have 40 dB isolation when terminated in 50 Ω although the isolation is reduced to 20 dB when they are terminated with open or short circuit impedances.

The test results for the preselector are shown in Figure 7.8.

| Channel Frequency (MHz) | Intermod Level Output | | | Preselector Gain (dB) |
	-1.5 & -3 MHz Input (dBm)	-3 & -6 MHz Input (dBm)	-10 & -20 MHz Input (dBm)	
240	-26	-68	-84	0
300	-2	-26	-88	+4.6
380	-24	-42	-88	-6.1

Figure 7.8   Intermodulation Test Results

H.   Tuning Speed

Tuning speed was measured using the test configuration shown in Figure 7.9. The RF level output of the preselector is measured with an RF detector. This signal is then monitored on one channel of an oscilloscope while the frequency change command is monitored on the other channel.

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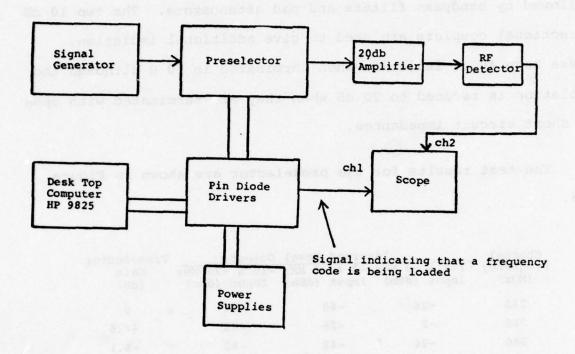
Figure 7.9 Preselector Tuning Speed
Measurement Test Setup

The tuning speed from the off to the on state is shown in Figure 7.10a. The tuning time of 3 μseconds, is limited by the carrier lifetime of the pin diodes. The tuning time for the other direction from the on to off state is shown in Figure 7.10b. This is somewhat slower, approximately 400 μseconds, since a simple pull up resistor is used for the reverse bias condition to reduce the complexity of the tuning network. If extra transistors were added to provide active pull up, then both the on and off transistions would be approximately 3 μseconds.

I.   Size and Weight

The preselector filters and amplifiers occupy 80 cubic inches. This is very close to the size goal. The preselector weighs approximately four pounds which is double the weight goal.

It was decided to use fairly thick wall aluminum construction to aid in assembly and disassembly for test purposes. If thinner wall aluminum is used for follow on or production efforts, the weight goal can be met.
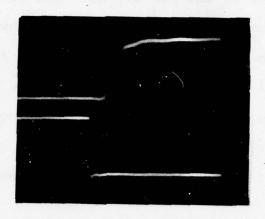
Figure 7.10a   Tuning Speed from Diode
Off to On (2us/div)
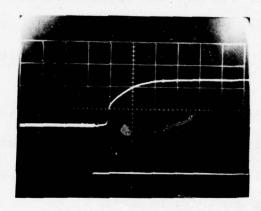
Figure 7.10b   Tuning Speed from Diode
On to Off (200us/div)

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VIII CONCLUSIONS

Small high dynamic range filters have been successfully demonstrated in a UHF preselector which tunes the entire 225 to 400 MHz frequency range.

The bandwidth and intercept point were substantially improved over earlier varactor tuned type filters. A new technique of tapping PIN diodes at a single point on the resonators and using PIN diodes for bias isolation gave excellent resonator tracking versus frequency.

A problem of maintaining constant insertion loss versus frequency was encountered and determined to be caused by resonator coupling changes versus frequency. This was partially solved by using short input and output transformers which had low coupling in the vicinity of PIN diode tuning networks although additional work to further reduce this problem is suggested.

IX SUGGESTIONS FOR FUTURE WORK

A. **General**

During the course of this investigation several ideas
have occurred which may warrant future study. They all con-
cern improving the performance of pin diode switched filters.

B. **Improved Pin Diodes**

The pin diodes are the major contributor to the filter's
losses. Diodes can be tailored to and optimized for low loss
in filter structures operating in the 225 to 400 MHz frequency
range. Current diode data indicates that these filters'
insertion loss can be reduced by a factor of two with better
diode design and packaging.

The diodes can also be optimized for lower on resistance
at lower currents which would lower the power dissipation for
diode switched filters and allow use in battery operated
manpack type radios.

C. **Improved Coupling Techniques**

Changes in coupling versus frequency caused insertion
loss variation and frequency tracking error not only for this
contract but also for those listed in references one and two.
This problem is proportional to the bandwidth and frequency

range which a pin diode switched filter must tune. Investigations should be performed into structures which minimize this effect. The position of coupling elements can be optimized and variable coupling networks using pin diode switched capacitors can be investigated.

D.   Application to a Satellite Band Receiver Preselector

The insertion loss of an electronically tunable filter is proportional to the tuning range over which it is tuned. The bandwidth can be narrowed or the insertion loss lowered, or both, if the filter is designed to tune a smaller portion of the UHF band. Satellite receiver preselectors require filters with insertion losses in the 1 to 2 dB range which can be achieved provided the filter's tuning range is limited to the 10 to 30 MHz band of downlink signals of typical UHF satellites.

The filter also could be designed to impedance match a low noise amplifier which could be mounted internally in the filter.

E.   Preselector Interface with a Higher Dynamic Range Amplifier

The high dynamic range amplifier currently being developed by RADC could be installed in the UHF preselector and performance data taken across the UHF frequency range.

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REFERENCES

1. A. Karp and W. B. Weir, "Electronically Tunable High-Power Filter for Interference Reduction in Air Force Communications Systems", Final Report, Contract F30602-71-C-0255, June 1972. RADC-TR-72-196, AD#748 196.

2. A. Karp and L. N. Heynick, "UHF Electronically Tunable High Power Filter", Final Report, Contract F30602-74-C-0142, September 1975. RADC-TR-75-220, AD#A018 044.

3. A. Karp, "Electronically Tunable UHF High-Power Filter", Final Report, Contract F30602-76-C-0248, May 1977.*

4. H. J. Blinchikoff and A. I. Zverev, Filtering in the Time and Frequency Domains, John Wiley and Sons, New York, 1976.

5. G. L. Matthaei, L. Young, E. M. T. Jones, Microwave Filters, Impedance-Matching Networks, and Coupling Structures, McGraw Hill Book Company, Inc., New York, 1964.

6. J. Getsinger, "Coupled Rectangular Bars Between Parallel Plates", IRE Transactions, PGMTT-10, pp. 65-72, January 1962.

7. H. H. Skilling, Electric Transmission Lines, McGraw Hill Book Co., New York, 1951.

*RADC-TR-77-167, AD#A040 469.

Appendix A

COMPUTER AIDED COMB LINE FILTER DESIGN PROGRAM

Computer aided design programs make the design of comb line filters to a set of dimension limitations much easier. Using the final equations for a two pole comb line filter, the program calculates the filter dimensions for a Hewlett Packard HP-9825 desk top computer. After the calculations are complete, the dimensions are used to plot the end view of the filter on an HP-9871 printer. Figure A.1 is a listing of the program written in HPL and Figure A.2 is an example output.

The concept of the program is to prompt the operator and request the required information to calculate the required capacitance per unit length values for the filter elements. The designer will input:

Resonator Series Resonate Frequency (MHz) = F_{sr}

Operating Frequency of Filter (MHz)   = f_o

3 dB Bandwidth at Operating
    Frequency (MHz)   = BW

Input Impedance (ohm)   = Z (input and
                                           output)

Resonator's Impedance (ohm)   = Z resonator

Cavity Internal Height (inch)   = b

Element Thickness (inch)   = t

After the per unit length capacitance for the elements are found, the program prompts the designer for input from two graphs to find the element to element spacing and then solves for element widths. These graphs are shown in Section III. The HP-9871 printer/plotter plots the dimensions to an accuracy of one-ninetieth of an inch.

```
0: "Comb Line Bandpass Filter Dimensions":
1: "6 Jan 78  Tape 000 Track 0 File 4":
2: cfg ;fmt 0,c,f7.2,c;rad
3: wrt 0,"Resonator Frequency ",A,"Mhz";ent "",A
4: wrt 0,"Operating Frequency ",B,"Mhz";ent "",B
5: wrt 0,"Bandwidth ",C,"Mhz";ent "",C
6: wrt C,"Input Impedance ",D,"ohm";ent "",D
7: wrt 0,"Resonator Impedance ",E,"ohm";ent "",E
8: wrt 0,"Cavity Internal Height",F;ent "",F
9: wrt 0,"Resonator Thickness ",G;ent "",G
10: πB/2A→Z; C/B→Y;G/F→P
11: fxd 4;prt "t/b=",P
12: (D/E)((1/tan(Z)+Z(1/sin(Z))↑2)/2)→X
13: YX/γ2→W
14: (376.7/D)(1-γW)→V
15: (.376.7/D)(D/E+W(1-tan(Z))-γW)→U
16: 376.7/D-V→T
17: (376.7/D)(Wtan(Z))→S
18: fmt 1,c,f7.4,c,f7.4
19: wrt .1,"C01=",T," Using 5.05-9 S01=",H;ent "",H
20: wrt .1,"S01=",H,"Using 5.05-9 C01=",I;ent "",I
21: wrt .1,"C12=",S," Using 5.05-9 S12=",J;ent "",J
22: wrt .1,"S12=",J," Using 5.05-9 Cfe12=",K;ent "",K
23: wrt .1,"t/b=",P,"Using 5.05-10 Cf=",L;ent "",L
24: .5(1-P)(.5V-I-L)→R
25: .5(1-P)(.5U-I-K)→Q
26: if R/(1-P)<.35; (.07(1-P)+R)/1.2→R;sfg 1
27: if Q/(1-P)<.35; (.07(1-P)+Q)/1.2→Q;sfg 2
28: if R/(1-P)<.1;sfg 3
29: if Q/(1-P)<.1;sfg 3
30: if flg3;ent "W/b is too small--try again",0;gto 0
31: 0→O;ent "Is everything OK? (Yes=CONT No=1)",O;if O;gto 0
32: wtb 6,10,13,10;fmt 2,c,f8.2,c,f8.2,z;fmt 3,c,f8.2,c
33: wrt 6.2,"Resonator Frequency ",A,"Mhz  Operating Frequency ",B
34: wrt 6.3,"Mhz  Bandwidth ",C,"Mhz"
35: fmt 4,c,f8.3,c,f8.3,c
36: wrt 6.4,"Input Impedance ",D,"ohm  Resonator Impedance ",E,"ohm"
37: fmt 5,c,f8.4,c,f8.4,c,f8.4
38: wrt 6.5,"t/b ratio ",P," w/b cf input ",R," w/b of resonators ",Q
39: wrt 6.5,"s/b between input and resonator ",H," s/b resonators ",J
40: wrt 6.5,"Thickness of Resonators and Input ",FP," Cavity Height ",F
41: wrt 6.5,"Width of Input ",RF," Width of Resonators ",QF
42: wrt 6.5,"Distance input and resonator",HF," Distance Resonators ",JF
43: wrt 6.5,"Total Length greater than ",((R+Q+H)2+JK)F
44: wrt 6.5,"Resonator length aproximately ",2808/A
45: 0→O;ent "Draw end view (Yes=CONT  No=1)",O;if O;gto 0
46: wtb 6,10,13,10,27,84,27,77
47: wtb 6,27
48: wtb 6,114,int(120(4H+(R+Q)2+J)F/64),int(120(4H+(R+Q)2+J)F),int(0),int(0)
49: wtb 6,27
50: wtb 6,82,int(-120(HF)/64),int(-120HF),int(-96(F-G)/128),int(-96(F-G)/2)
51: RF→N;G→O
52: gsb "SQ"
53: wtb 6,27,82,int(-120(HF+N)/64),int(-120(HF+N)),int(0),int(0)
54: QF→N;gsb "SQ"
55: wtb 6,27,82,int(-120(JF+N)/64),int(-120(JF+N)),0,0
56: gsb "SQ"
57: wtb 6,27,82,int(-120(HF+N)/64),int(-120(HF+N)),0,0
58: RF→N;gsb "SQ"
59: wtb 6,27,82
60: wtb 6,int(-120(HF+N)/64),int(-120(HF+N)),int(-96(G+F)/128)
61: wtb 6,int(-96(G+F)/2)
62: wtb 6,27,114,int(120(4H+(R+Q)2+J)F/64)
63: wtb 6,int(120(4H+(R+Q)2+J)F),0,0
64: wtb 6,10,13,10,10;gto 0
65: "SQ":
66: wtb 6,27,114,int(0),int(0),int(-96O/64),int(-96O)
67: wtb 6,27,114,int(-120N/64),int(-120N),int(O),int(0)
68: wtb 6,27,114,int(0),int(0),int(96O/64),int(96O)
69: wtb 6,27,114,int(120N/64),int(120N),int(0),int(0)
70: ret
71: end
72: dim A$[3]
73: ent A$
74: wrt 0,A$;ent "",r1
75: fmt 1,c3,2x,,f10.5;wrt 16.1,A$,r1
76: jmp -3
*29411
```

Figure A.1 Comb Line Filter Dimension Program

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Resonator Frequency 1230.00MHz   Operating Frequency 312.00MHz   Bandwidth 50MHz:
Input Impedance 50.000ohm   Resonator Impedance 77.000ohm
t/b ratio 0.4000 w/b of input 0.1375
s/b between input and resonator 0.5811 w/b of resonators 2.0200
Thickness of Resonators and Input 0.7500 s/b resonators 2.0200
Width of Input 0.5811 Width of resonators 1.0000
Distance input and resonator 0.7500 Cavity Height 0.1375
Total length greater than 5.0572 Distance Resonators 2.0200
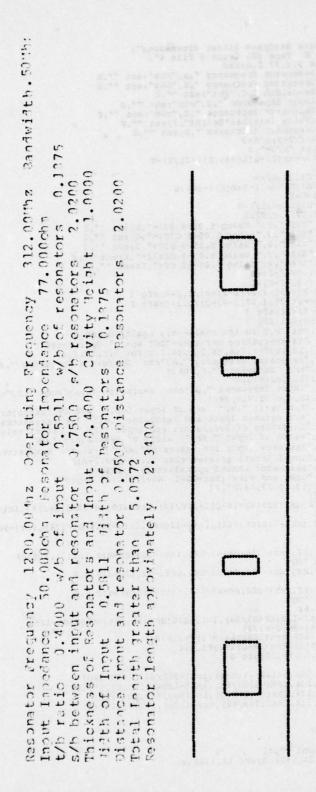Resonator length approximately 2.3400

Figure A.2   Comb Line Filter Example Output

Appendix B

SCHEME FOR INTERFACE TO ARC-164 OR ARC-171

Interfacing the electronically tuned preselector to the AN/ARC-164 or AN/ARC-171 is shown in Figure B.1. A microprocessor receives and interprets the serial frequency control data from radio control lines and allows interface with different types of frequency control systems. The 50 ohm input and output impedance of the preselector is standard for both the ARC-164 and ARC-171.

A read-only memory is used to store the tuning code information required for the setting of the frequency of the filter. The use of the ROM allows the system to be modified for different kinds of filters by changing the ROM and filter. The serial bit stream is available on connector J3 of the ARC-164. The microprocessor allows flexibility so that software changes are all that are necessary for operation with the ARC-171.
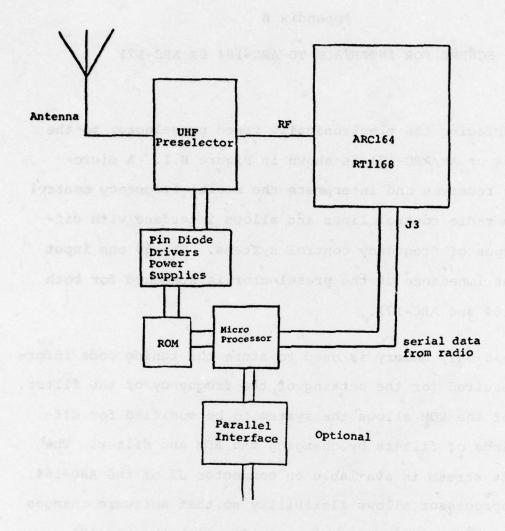
Figure B.1   Interface Scheme for
ARC-164 or ARC-171

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